Goddard

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UNIVERSITY OF MICHIGAN
Department of Astronomy

FINAL REPORT
Contract NASw-55

GENERAL

Contract NASw-55 was initiated on 29 April 1959, originally for a period of one year. The purpose of the contract was to carry out preliminary design studies looking toward the construction of two ultraviolet scanning spectrometers to be used for the observation of the solar spectrum on board the S-16 satellite. In addition to the Principal Investigator, Dr. Leo Goldberg, the major participants in the studies were Dr. William Liller, Principal Collaborator, Dr. James A. R. Samson, Research Physicist, Mr. Lewis W. Wolf, Mechanical Engineer, and Mr. William H. Follett, Electronics Engineer. At the end of January 1960, Drs. Goldberg and Liller decided to accept appointments to the faculty of Harvard University, beginning 1 July 1960, and since it was their wish to continue the work at Harvard, the University of Michigan decided not to request additional funds to extend the work beyond the preliminary design phase. However, no-cost extensions were requested and granted up to a final termination date of 31 October 1960, in order to permit the completion of the contract tasks. By mutual agreement, title to all non-expendable equipment purchased by Michigan under the contract was transferred to Harvard University, Contract NASw-184.

The main body of this report consists of two appendices describing work carried out in the design of the electronic system and of the mechanical-optical system of the proposed satellite instrument.

Leo Goldberg

Leo Goldberg
Principal Investigator

LG:hf

1 March 1962

Appendix 1

THE UNIVERSITY OF MICHIGAN -- WILLOW RUN LABORATORIES

ELECTRONIC SYSTEM FOR S-16 SATELLITE

by

William H. Follett

1. INTRODUCTION

Two spectrometers will be used to cover the spectral region 1500 A to 100 A; a normal-incidence instrument will cover the region from 1500 A to 500 A, and a grazing-incidence instrument the region from 600 A to 100 A.

The long-wave instrument uses a 70-cm radius of curvature grating ruled 30,000 lines per inch and an entrance slit 20 μ wide by 100 μ high. The width of the exit slit will be interchangeable between 20 μ and 100 μ, depending on whether a complete spectral scan is carried out in 20 or in 4 minutes. The fast scan speed will be used on fast-changing phenomena such as solar flares, while the slow scan will provide high-resolution spectra of normal regions. The scanning will be accomplished by rotating the grating on an offset pivot. The short-wave instrument is similar to the long-wave instrument except that it will use a 1-meter concave grating and the exit slit will be moved along the Rowland circle during scanning.

The electronics for both spectrometers are identical, although the resolution of the normal-incidence spectrometer is somewhat better than that of the short-wave instrument. Therefore, the electronic system was designed to fit the requirements of the normal-incidence instrument.

The computed spectral resolution of the long-wave spectrometer is 0.24 A. Since the spectral range is 1000 A, there are 4170 (rounded off to 4000) pieces of information per scan, which are required to be reproduced on the ground within one percent error. Such high accuracy can best be achieved by digital methods.

The method chosen to measure spectral intensity was to count incoming photons with a gated binary counter. The incoming photons incident on a photocathode are multiplied by 10^7 in an electron multiplier. The output of the photomultiplier is therefore a burst of 10^7 electrons per incident photon. Each burst produces a voltage pulse that is amplified and fed into a 15-stage binary counter. The output of the counter is a 15-bit binary word in serial form. The gate on the binary counter allows the counter to sample each of the 4000 spectral pieces three times. The total number of electrical samples is then 12,000.

During fast scanning, both the slit width and scan speed are increased by a factor of five, but the sampling rate is unchanged; therefore, the number of samples per slit width remains at three.

At both slow and fast scan speeds, the sampling rate is 10 per sec. Since the sample is stored in a 15-stage counter, the output of each of the 15 stages of this counter represents the intensity of the sample by a 15-bit binary code. Since these 15 bits represent one sample or one intensity measurement, they are called a <u>word</u>. The information output of one spectrometer is then ten 15-bit binary words per sec, or 150 bits per sec.

This information is transmitted directly when the satellite is in contact with a ground station, and is stored on tape at other times. If the satellite is in sunlight for 60 minutes per orbit, the tape recorder must store 150 x 60 x 60 = 540,000 bits of information per spectrometer. Information packing density on tape approaches 1000 bits per track per inch of tape. The minimum tape length needed is then 540 inches or 45 feet. It would be advisable to use at least 70 feet of tape because as the packing densities get high the recorder becomes quite critical in adjustment.

The scan drive motor is a stepping motor with magnetic detents. The motor requires an 87-ma pulse to develop full torque output. The motor is then held by the detent until a second pulse of opposite polarity is applied to the motor.

The motor will drive a gear box that in turn will rotate the grating of the long-wave instrument or slide the exit slit along the Rowland circle in the grazing instrument. It is supposed that the gear box will filter the mechanical steps, and therefore the motor is pulsed at half the rate of the scan oscillator.

The remainder of this report will describe and discuss the design of the various units used to perform the tasks outlined above.

2. PHOTON COUNTING

Photon counting is the most sensitive and accurate system for measuring light intensity. It is sensitive because current not due to incoming photons can be rejected by means of electronic circuitry. In current-measuring systems, pulses due to electrons emitted at points other than the photocathode, leakage current and amplifier noise are all added to the output current and therefore limit sensitivity. In photon counting, discrimination between the current due to other causes and that due to incoming photons greatly improves sensitivity. The limit of sensitivity in the photon-counting system is set by the rate of pulses due to incoming ions rather than photons. As to accuracy, the photon-counting system is inherently perfect since the method counts the number of incoming light quanta (actually, it counts a certain percentage of the incoming quanta, 10 percent in the case of tungsten). The measurement can be made to any desired accuracy by simply increasing the time of observation and therefore increasing the total number of counts. Other advantages of photon counting are as follows. (1) The gain of the photomultiplier and amplifier affects only the pulse height and not the pulse rates. (2) The ultimate dynamic range is limited by the noise, which is less than 1 pulse per sec in the Bendix photomultiplier, to nearly 10^8 pulses per sec, the point at which the pulse width becomes wider than the period of the pulses. This range can be covered without any adjustment or change of scale.

The criterion for setting the gain of the photomultiplier is to assure that the pulse height is sufficiently above the noise of the pulse amplifier to assure detection of the output pulse.

For most purposes (counting rates less than 10 mc), the output pulse height from the photomultiplier is determined by the capacitor across the photomultiplier load resistor and the gain of the photomultiplier. In most photomultipliers, the difference in arrival time of the electrons in a given burst is so small that all the electrons essentially charge the capacitor instantaneously. They then bleed off according to the product of the load R and the capacitance C. The pulse height is then

$$C_o = \frac{Gq}{C} ,$$

where G is the gain of the photomultiplier; q is the electronic charge; and C is the capacitance across the load resistor. Since the output voltage of this network is inversely proportional to the capacitance, keeping the capacitance low will reduce the required gain of the video amplifier. A minimum gain for the photomultiplier can be determined by setting the output pulse height equal to the noise of the circuit to which the photomultiplier is connected. The noise of such a circuit is

$$e_n^2 = 4KTBR ,$$

where K is Boltzmann's constant; T is the absolute temperature; B is the noise bandwidth; and R is the circuit resistance. The noise bandwidth of an RC network is the value at the 3 db point, or

$$B = \frac{1}{2\pi RC} .$$

Therefore,

$$e_n = \sqrt{\frac{2KT}{\pi C}} .$$

If we equate e_n to C_o, we obtain

$$G = \sqrt{\frac{2KTC}{\pi q^2}} .$$

However, since noise is made up of random pulses, the gain of the photomultiplier must be increased until the probability that a noise pulse will equal the signal pulse is sufficiently small. This requires an additional factor of something like 4.7 in order that the noise give about 100 spurious counts per sec of the designed 10-mc bandwidth. This then gives one spurious count per sampling period. If the input capacitance were 10 μμf, the required gain to satisfy the above criterion would be 1.47×10^4, which is easily obtainable with nearly all photomultipliers.

3. VIDEO AMPLIFIER

The video amplifier widens the pulse to a width needed to trigger the binary counter and raises the input pulse level to that needed to trigger the counter.

Widening is accomplished by the input circuit by controlling the discharge time constant of the input capacitor. This value will be about 3×10^{-8} sec (the exact value will be specified by Strand Engineering, who are building the binary counter).

As previously mentioned, the gain required by the amplifier is inversely proportional to the input capacitance; therefore, effort was put into developing a low capacitance input stage. A two-stage emitter-follower preamp is used as an input circuit as shown in Fig. 1; the input circuit capacitance achieved was about 1 $\mu\mu$f.

In an emitter-follower circuit, the input capacitance is

$$C = C_o/\beta + C_c + C_s \ , \qquad (1)$$

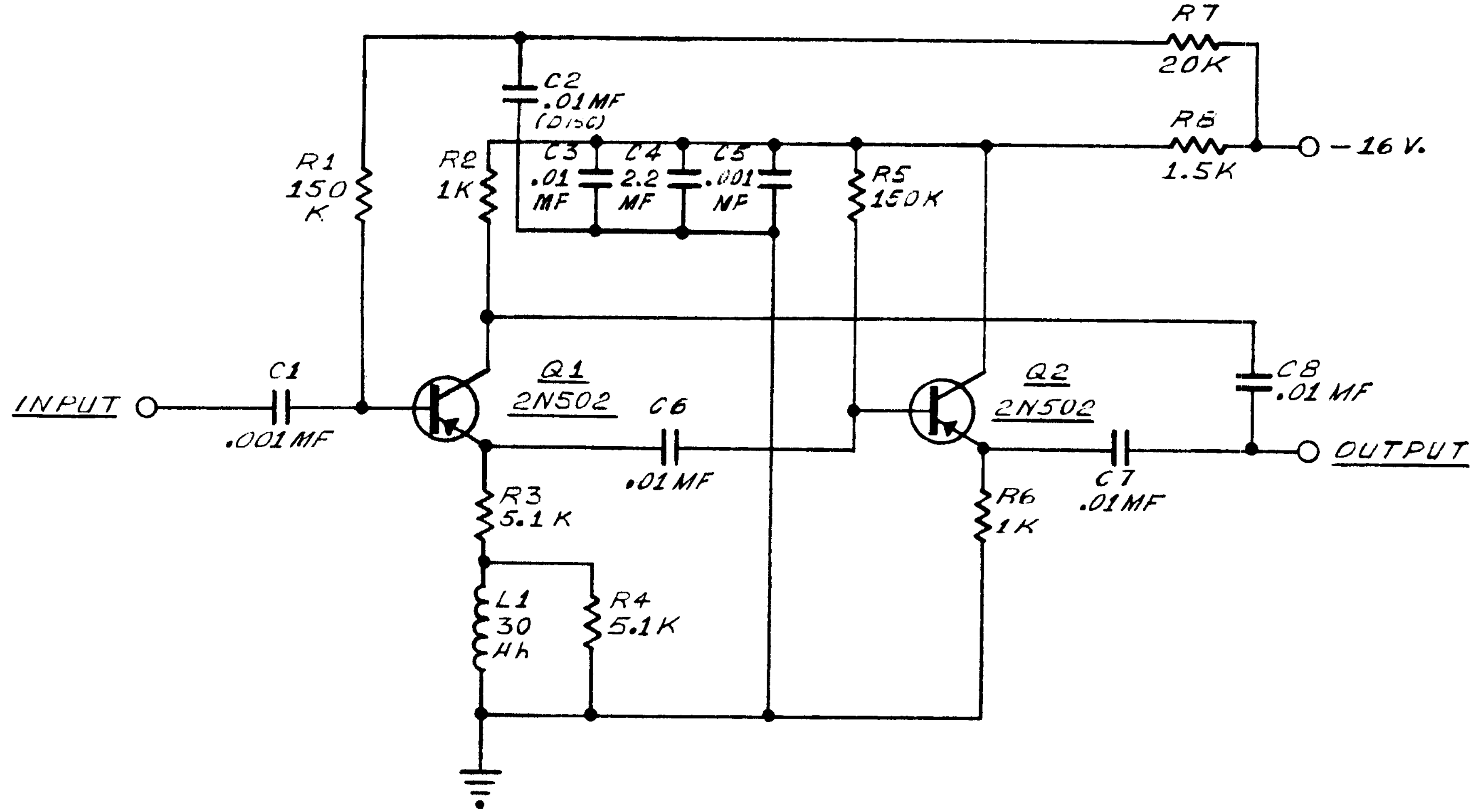
where C_o is the output load capacitance (input capacitance of main amplifier); β is the base current gain of the transistor; C_c is the base-to-collector capacitance; and C_s is the stray capacitance due to wires, etc.

In the emitter-follower design, C_o was reduced by a second emitter-follower, thereby reducing the first term to C_o/β^2.

The second term was reduced by making the AC voltage of the collector nearly equal to the input voltage, thus reducing the effective capacitor. The amount of reduction is proportional to the amount by which the collector

Figure 1
Page 7a

K&E 195T 2-58



UNLESS OTHERWISE SPECIFIED, STANDARDS ARE:
RESISTORS: ½ WATT CARBON ± 10%
CONDENSERS: PAPER 600 V.D.C.
MICA & CERAMIC 500 V.D.C.

POT.: TYPE JU
PLAIN: 3/8 BUSHING 3/4 STRAIGHT SHAFT
SLOTTED: 3/8 BUSHING 1/2 SLOTTED SHAFT

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DR.	De Veé	8-19-60

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UNIV. OF MICHIGAN
WILLOW RUN

SCHEMATIC-
VIDEO AMPLIFIER
PREAMP

J-61627
02933 D-005

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ISSUE

REVISED

X4800

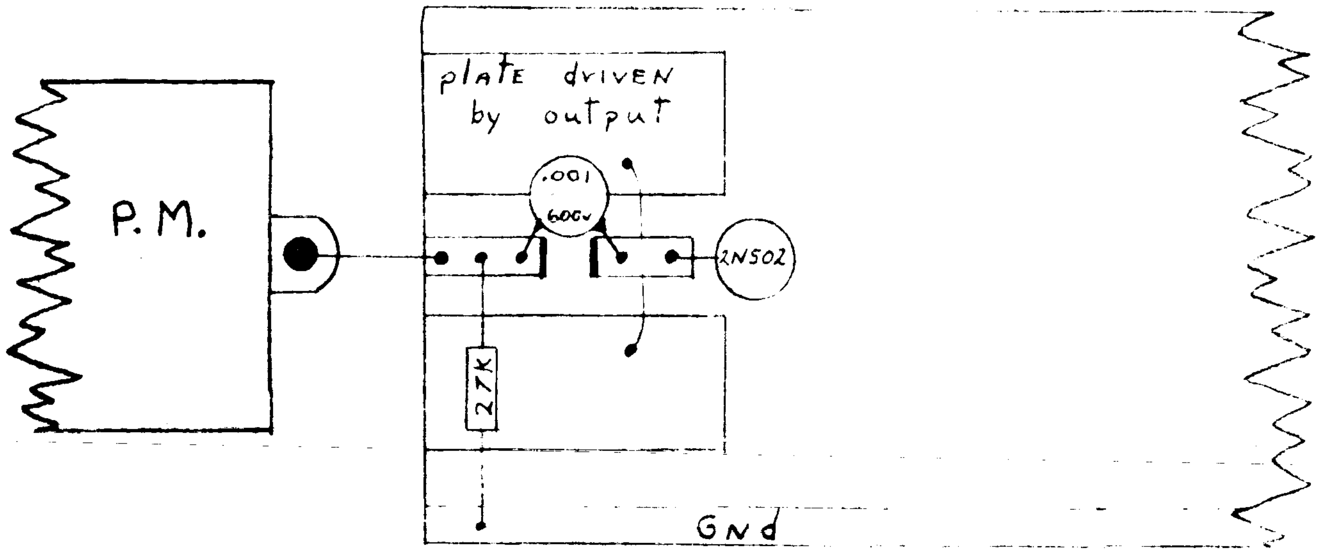
this energy reaches the preamp, and that it is not radiated to other components or impeded by lead inductance. A drawing of a suitable connection is shown in Fig. 2.

The output impedance of the preamp is quite low and will drive several hundred micromicrofarads. However, care must be taken that the grounded metal frame of the spectrometer is not placed too close to the preamp shield.

The output of the preamp is fed to the main amplifier by a coaxial cable. The length of this cable should be such that the capacitance due to the cable is less than 50 $\mu\mu$f.

The main amplifier is a published circuit found in a Government Printing Office publication entitled Handbook of Selected Semiconductor Circuits. The circuit was modified to use 16 volts instead of the 27 volts for which it was designed. Since the DC emitter current is the primary factor that determines gain, it was kept at the former design value. This resulted in a lowering of the collector-to-emitter voltage, which affects transistor gain very little. The lowering of the maximum output voltage obtainable is also of no concern since the output voltage required is well under the new maximum. The collector resistor R_{12} was lowered to 1 K to provide a load on this transistor since the output of this transistor was fed into an emitter-follower.

The bandwidth of the main amplifier is adjustable with a maximum of nearly 20 mc, the gain bandwidth product being about 1300. The bandwidth of the amplifier is to be adjusted so that the width of the input pulse is not widened greatly. This bandwidth will be from 12 mc to 15 mc. Appreciable



If possible, operate PM with anode at ground i.e., input end of dynode strip at -2000 volts.

Figure 2

change in the pulse width would indicate a loss in the available gain of the amplifier. The reason, of course, is that a widening of the pulse would indicate that the upper frequencies of the pulse spectrum were not being passed by the amplifier. Widening of the pulse should be done only in the input circuit.

The amplifier will be allowed to change the pulse shape. The rise time of the pulse at the photomultiplier is about 10^{-8} sec. This can be slowed to nearly 3×10^{-8} sec without serious loss in available gain. Thus the input pulse to the preamp will have very fast rise with an RC decay of about 6×10^{-8} sec, while the output pulse from the amplifier will have more nearly equal rise and fall times, but no greater pulse width.

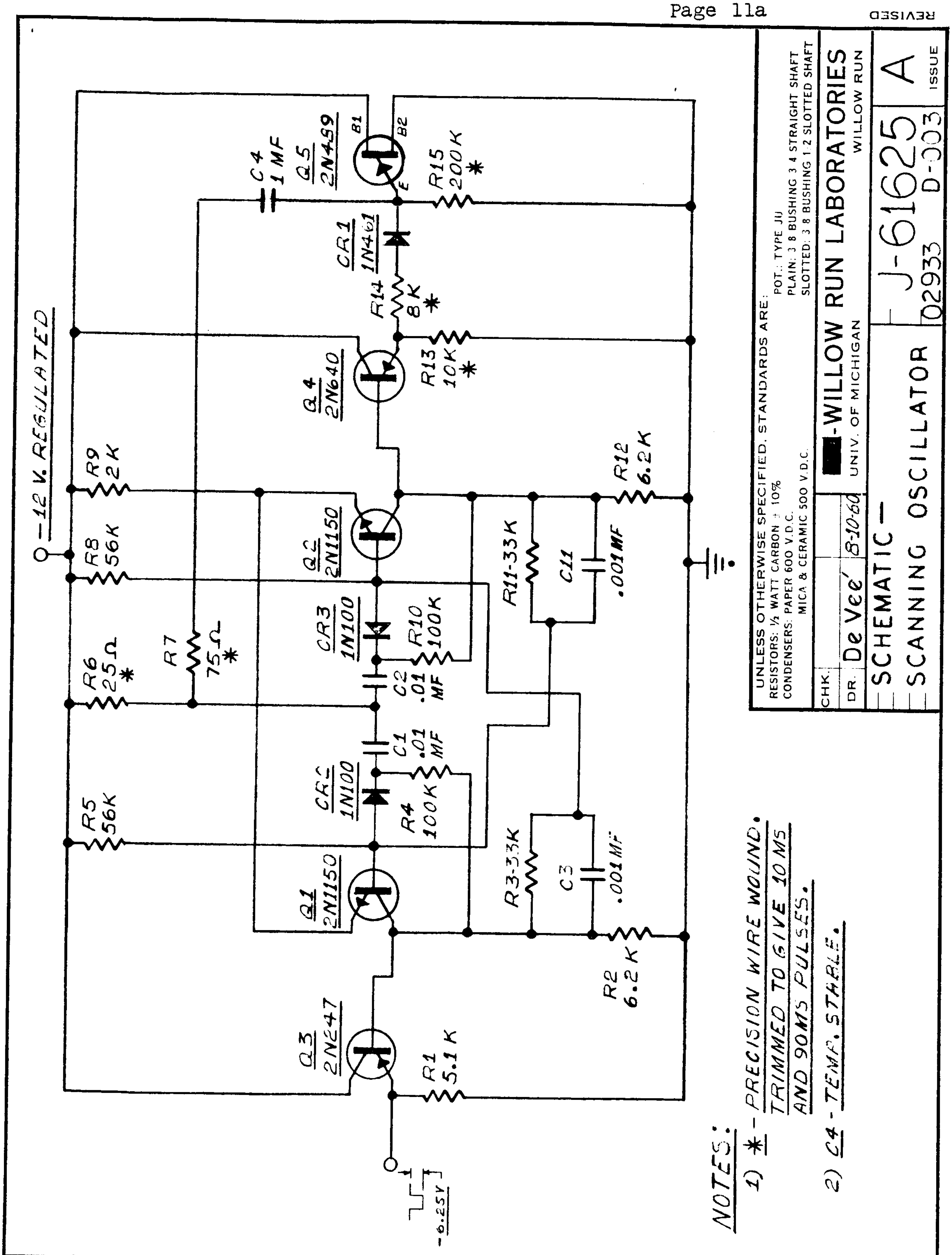
4. SCAN OSCILLATOR

The scan oscillator provides the basic timing in the system. It provides both a 10-ms pulse to operate the gate of the binary counter, and the sample timing of 100 ms. This is accomplished by use of a free running multivibrator whose basic period is the sampling time period, i.e., 100 ms. The output is a 10-ms pulse separated by 90 ms.

In order that the system be capable of measuring intensity with one-percent accuracy, the binary-counter gate period must have a long time stability better than 0.3 percent (this is an arbitrary figure that assumes there will be other errors and that the gate-period uncertainty should not contribute appreciably to the other errors). Since normal transistorized free-running multivibrators will not give this stability, a General Electric unijunction transistor was used as the basic timing element. This transistor works like a thyratron. When the emitter voltage reaches a certain value, the emitter-base-2 junction becomes a very low impedance. Figure 3 shows the circuit used. Timing is accomplished by charging C_T through R_{15} for the 90-ms period, and through R_{13}, R_{14}, and R_{15} for the 10-ms period.

When the unijunction fires, it produces a pulse across R_6 that flips the bistable circuit made up of q_1 and q_2. When q_2 is conducting, its collector voltage holds the emitter of q_4 at a more negative voltage than the emitter of q_5, the unijunction transistor. This backward biases CR_1, and therefore C_4 is charged only through R_{15} for a 90-ms period. Upon firing, the collector voltage of q_2 is near 0; therefore, C_4 now charges through R_{13}, R_{14}, and R_{15} for a 10-ms period.

Figure 3
Page 11a



The sum of R_6 and R_7 forms a stabilizing resistor to offset the drift of $Q_5 + CR_1$ against temperature, thus stabilizing the 10-ms pulse. The unijunction transistor is isolated from the bistable circuit by the voltage divider R_6 and R_7. Without this divider there was some feedback that affected the stability of the 10-ms pulse and caused excessive period change with temperature.

CR_2, CR_3, R_4 and R_{10} provide a steering circuit to ensure that the bistable circuit flips on every pulse. Q_3 provides the scanning oscillator with a low-impedance output, and Q_4 prevents loading of the bistable circuit when the timing circuit is charging through R_{14} and R_{13}.

The circuit is quite insensitive to supply voltage changes. A change in supply voltage of 1 volt changes the 10-ms period by about 0.2 percent. Zener diodes therefore should provide sufficient regulation of the voltage.

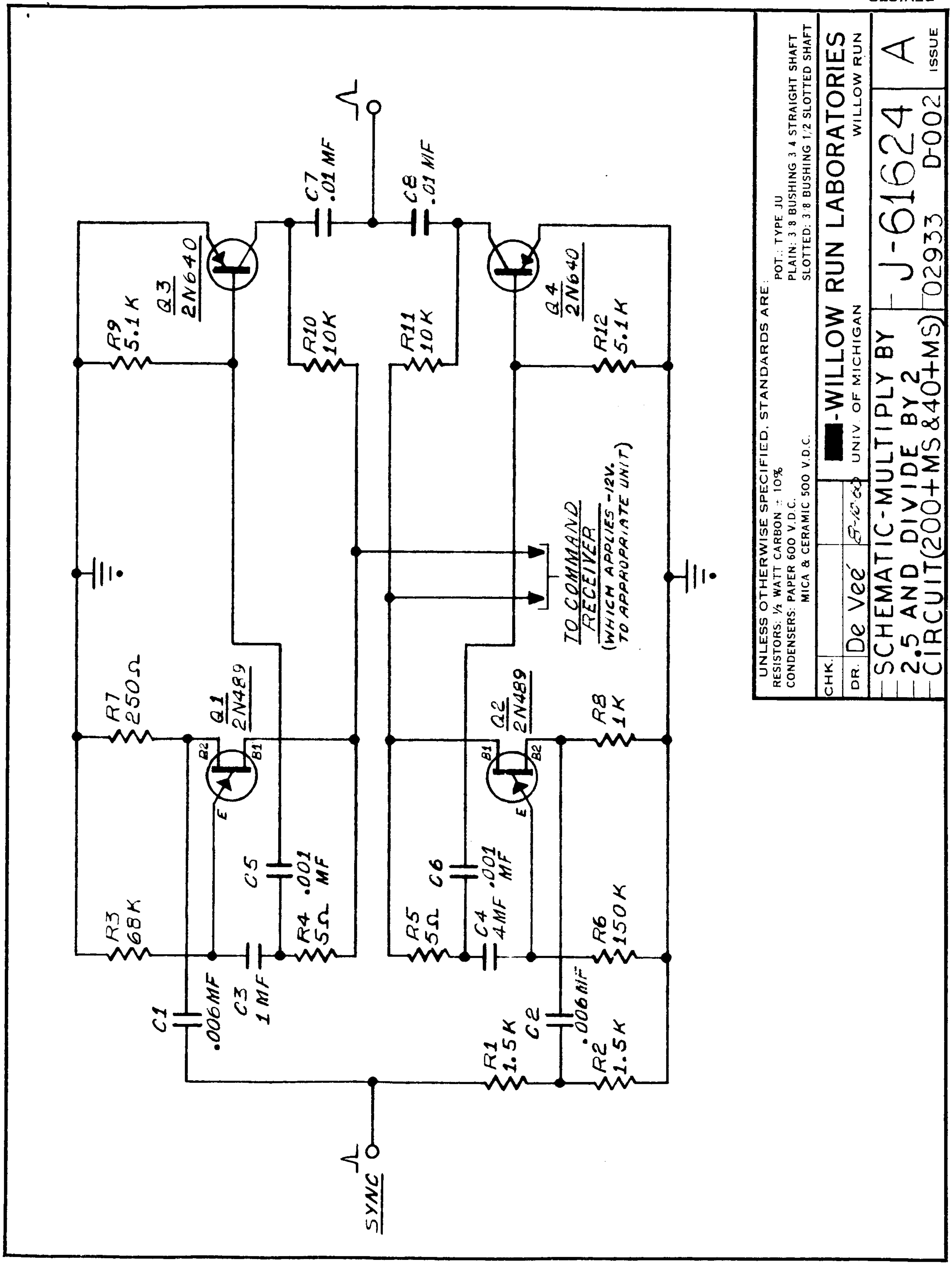
5. STEPPING MOTOR DRIVE

The stepping motor is driven by current pulses at 2 times the basic scan-oscillator period, i.e., 200 ms for the slow scan speed and 0.4 times the scan-oscillator period, i.e., 40 ms for the fast scan speed. One or the other of these pulse rates is chosen by a command signal from the ground, and fed to a pulse-shaping network. The shaped pulses then drive the output transistors, which in turn drive the stepping motor. The circuits are push-pull because the stepping motor requires a flux-reversal on successive steps. This flux-reversal is accomplished by means of a tapped motor winding (actually there are two separate windings but the two are tied together to make a centertapped winding). The driving circuit then puts a pulse first in one end of this winding, and a successive pulse in the other end of the winding, which requires a push-pull driving circuit.

5.1 MULTIPLY-DIVIDE CIRCUIT
(see Fig. 4)

The multiply-divide circuit consists of two free-running multivibrator circuits. The multiply circuit is set to free run at 25 pulses per sec (40-ms period) less about 2 percent. On every 5th pulse the multivibrator is triggered by the scan oscillator; this keeps the multivibrator's output synchronized to the scan oscillator.

The divide circuit free runs at 5 pulses per sec (200-ms period) less about 2 percent. This circuit is triggered by every other pulse from the scan oscillator, i.e., every multivibrator pulse, and thus is also synchronized to the scan oscillator.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UNIV. OF MICHIGAN
WILLOW RUN
CHK.
DR. De Veé 8-10-62
SCHEMATIC-MULTIPLY BY 2.5 AND DIVIDE BY 2 CIRCUIT(200+MS&40+MS)
J-61624
A
ISSUE
02933
D-002
X4800
SYNC
C1 .006MF
C3 1MF
R3 68K
R4 5Ω
C5 .001 MF
R7 250Ω
Q1 2N489
B2
B1
E
R1 1.5K
R2 1.5K
C2 .006MF
R6 150K
C4 4MF
R5 5Ω
C6 .001 MF
Q2 2N489
R8 1K
TO COMMAND RECEIVER
(WHICH APPLIES -12V. TO APPROPRIATE UNIT)
R9 5.1K
Q3 2N640
R10 10K
R11 10K
Q4 2N640
R12 5.1K
C7 .01 MF
C8 .01 MF

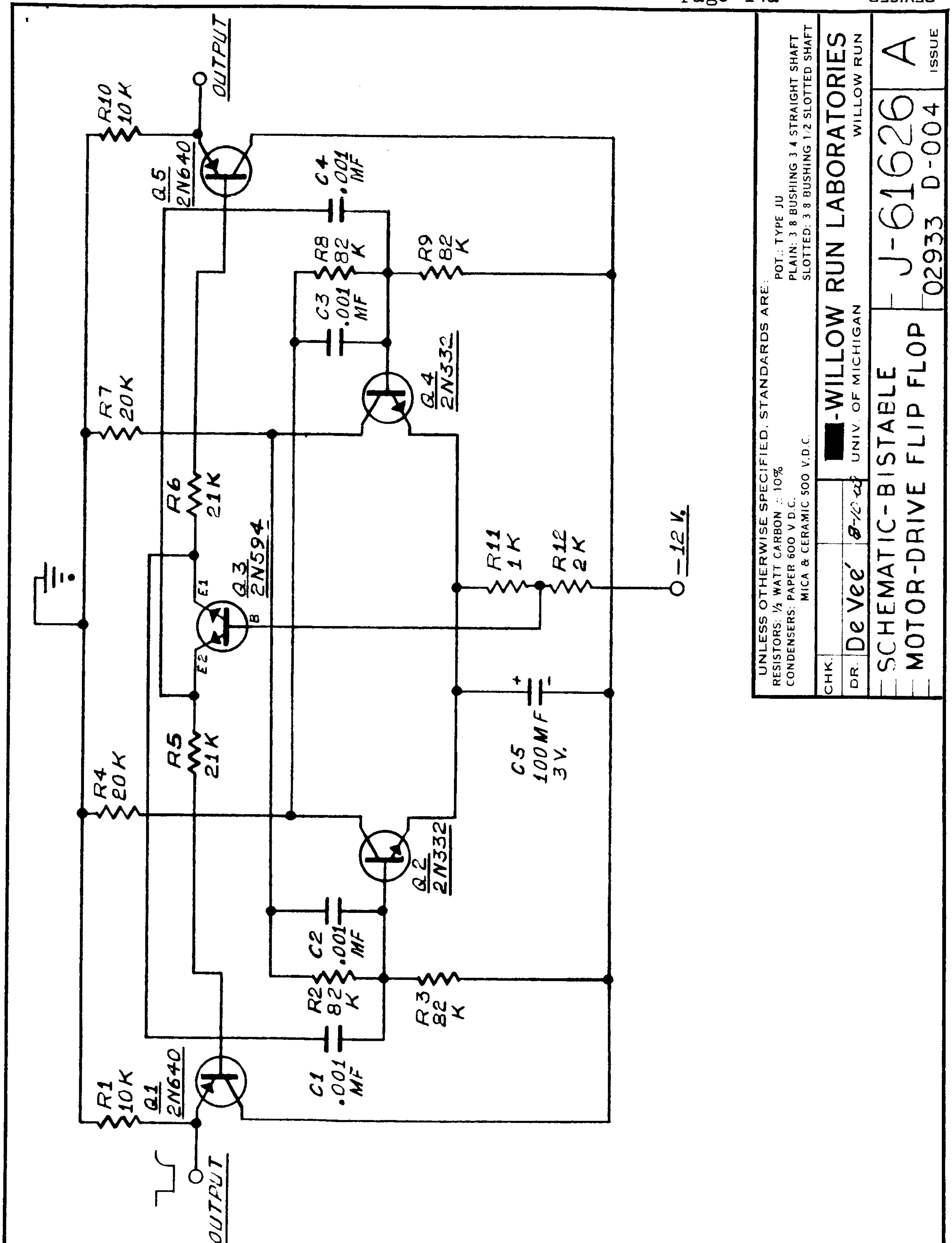
The multivibrator circuits use unijunction transistors in simple relaxation oscillator circuits with an emitter-follower output. The sync signal from the scan oscillator lowers the base-2 voltage, thereby reducing the trigger point to less than the emitter voltage, which causes the circuit to fire.

Speed switching is accomplished by allowing the satellite command receiver to apply supply voltage to one or the other of these circuits.

The output of the multiply-divide circuit is pulsed with the proper period, i.e., 40 ms or 200 ms. These pulses flip the motor drive flip-flop, which has a push-pull square-wave output (see Fig. 5). The circuit is a common bistable circuit but with a bilateral transistor used for pulse steering (Design of Transistorized Circuits for Digital Computers by A. Pressman, John F. Rider, New York, 1959, p. 11-298).

5.2 PULSE-SHAPING CIRCUIT (see Fig. 6)

The stepping motor requires a current of 37-ma peak to overcome the detent forces. The current between this level and the core saturation level (86 ma) develops torque to overcome the load inertia. The pulse duration or width also determines the torque output of the motor. A trade-off between pulse amplitude and pulse duration can be used to help match the motor to the load. A high-inertia load would more efficiently use a low-amplitude long-duration pulse, whereas a high-friction load would best be driven by a high-amplitude pulse of short duration, because a greater amount of the pulse energy would be used in developing output torque.



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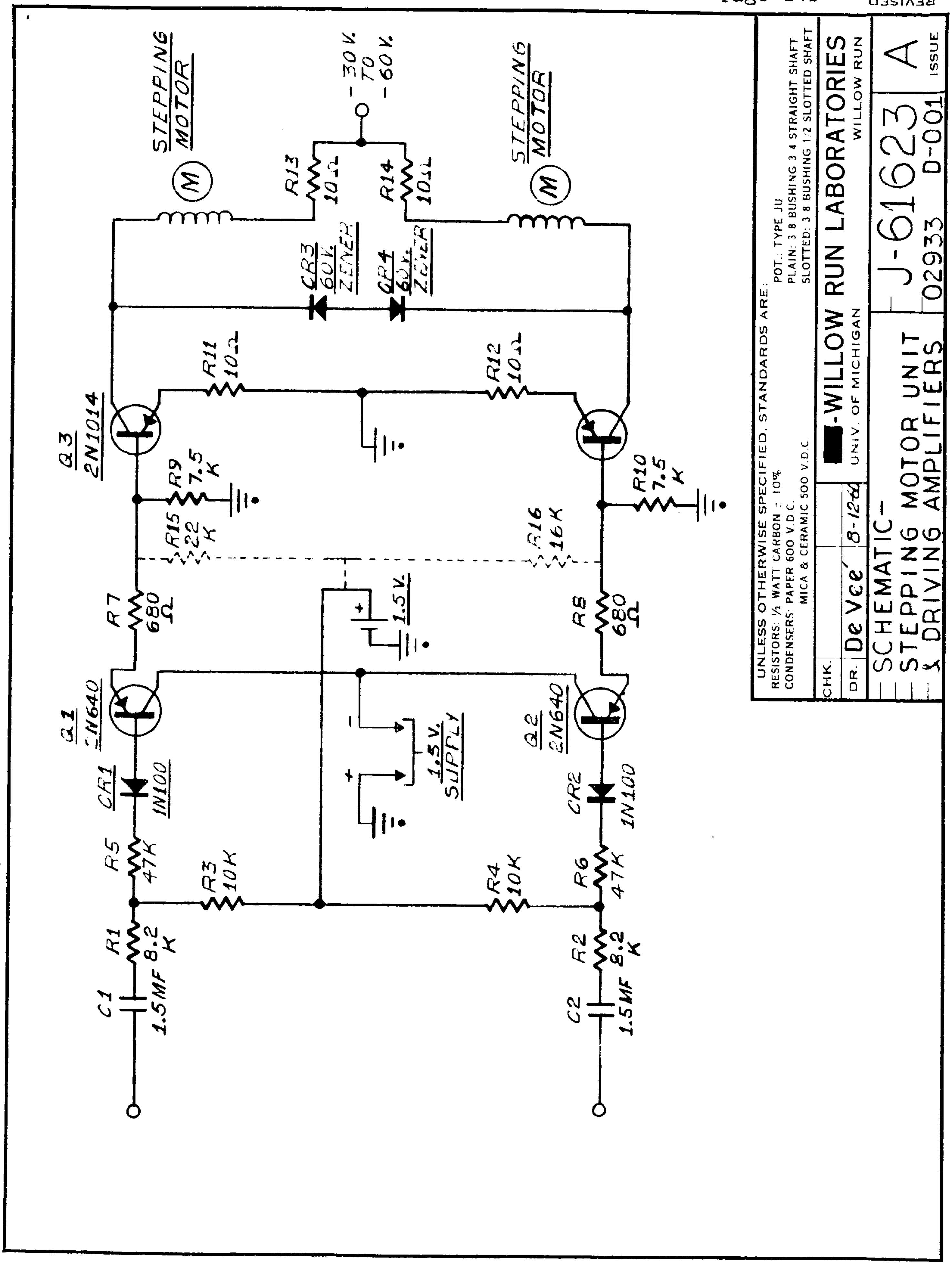
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CHK.
DR. De Vee' 8-12-60
WILLOW RUN LABORATORIES
UNIV. OF MICHIGAN
WILLOW RUN
SCHEMATIC-
STEPPING MOTOR UNIT
& DRIVING AMPLIFIERS
J-61623
A
ISSUE
02933
D-001
X-4800
STEPPING MOTOR
- 30 V. TO - 60 V.
R13 10Ω
R14 10Ω
CR3 60V. ZENER
CR4 60V. ZENER
Q3 2N1014
R11 10Ω
R12 10Ω
R9 7.5 K
R10 7.5 K
R15 22 K
R16 16 K
R7 680 Ω
R8 680 Ω
1.5 V.
Q1 2N640
Q2 2N640
1.5 V. SUPPLY
CR1 1N100
CR2 1N100
R5 47K
R6 47K
R3 10K
R4 10K
R1 8.2 K
R2 8.2 K
C1 1.5MF
C2 1.5MF

The rise of the current pulse is controlled by the motor R/L (14 ms), since the transistor is essentially a short circuit, i.e., it has a saturation resistance of the order of 1 Ω. The rise time to the desired current level is controlled by the supply voltage to the motor.

The height of the pulse is controlled by the 2N1014 driver transistor. When the current in the driver transistor reaches a certain value, the voltage across the transistor increases to the voltage necessary to keep the collector current at this value. This value of current is set by a resistor in the base circuit of the driver transistor. During the output pulse, the resistor is supplied by 1.5 volts. Thus the base current is limited to a value determined by the resistor.

The fall and therefore the width of the pulse is controlled by the time constant of C_1 $(R_1 + R_3)$. The input to this network is a -12 volt square wave supplied by the motor drive flip-flop. The output therefore decays from -12 volts to +1.5 volts. The network is returned to -1.5 volts instead of to ground in order to speed up the return of the voltage to zero. The network output voltage is impressed through base resistor R_3 to the base of Q_1. This resistor limits the base current when the base voltage exceeds the collector voltage. As long as this voltage is greater than 1.5 volts, the base current of Q_3 fed through Q_1 and R_7 is $1.5/R_7$. The emitter voltage will not rise above the collector voltage if the base current is limited; when the base voltage is above the collector voltage, the transistor voltage drop is only 0.1 volt. When the voltage applied to R_3 drops below 1.5 volts, the emitter current of Q_1 becomes proportional to the base current. Therefore, the collector current of Q_3 is proportional to the voltage impressed on R_3. On a consecutive pulse, the other half of the circuit operates as described above. Diode CR_1 keeps the positive peak from breaking down the base-emitter diode of Q_1 when the opposite side is operating.

As designed, the pulse height never reaches the saturation level because C_1 $(R_1 + R_3)$ is too short for the particular value of motor supply voltage used. Since the motor delivers more than adequate torque, resistor R_7 could be increased in value in order to limit the peak current to the motor to a lower value. Reducing C_1, R_1 and R_3 would also reduce the pulse height, since it is not now limited by resistor R_7, by cutting the pulse off quicker on the L/R rise curve. Since at high scan speed, the pulses run into each other, the motor supply voltage should not be reduced as a method of reducing output torque. This would reduce the pulse rise time, thus resulting in even greater pulse overlapping. The adjustment of output torque should be made by adjustment of C_1, R_1, R_3 and R_7 until adequate torque is delivered at minimum power at the fast scan speed. Of course, the resistors on the opposite half of the circuit should be changed to the same value. Since, as designed, the fast scan speed draws 625 mw and the slow scan speed draws 100 mw, conservation of power is more important at the high scan speed even though this scan speed will probably not be used as often as the slow scan speed. Also, power reduction at slow scan speed will be nearly proportional to the reduction at high scan speed.

It might be mentioned that at high temperature (above 30° C) the leakage of the transistors may deteriorate the performance slightly. For lab tests and demonstrations of output torque, backward-biasing resistors were used to reduce this leakage (shown dotted on schematic). These resistors were connected to a +1.5-volt battery and fed to the bases of Q_3 and Q_4 to keep the collector currents in the vicinity of 300 μa, at the higher operating temperature.

5.3 MOTOR DRIVING CIRCUIT

Since the driver circuit is push-pull, the collector voltage ranges from zero to twice the supply voltage (due to the emf developed when the opposite transistor starts conducting). This requires transistors that have a collector breakdown voltage of twice the supply voltage. Since at least a 30-volt supply was necessary, the collector voltage rating had to be in excess of 60 volts.

Zener diodes were placed collector-to-collector to ensure that the voltage did not exceed the collector voltage of the transistors. Since one collector voltage will be maximum when the opposite collector voltage is near zero, the Zener point chosen is slightly less than the collector voltage rating. Thus the Zeners will conduct when one collector is at a potential of the breakdown of the Zeners plus the collector of the opposite transistor.

6. BINARY COUNTER

The gated 15-stage binary counter is driven by the video amplifier and is capable of counting at 10-mc rates. The read-out is in serial form and is accomplished by converting the counter to a shift register.

The scan oscillator's 10-ms negative pulse opens the counter gate. The open gate admits pulses from the output of the video amplifier and thereby allows the counter to register the total number of pulses occurring in this 10-ms period. At the end of the 10-ms pulse, the gate closes and initiates a read sequence in the counter. The sequence starts by converting the counter into a shift register. The shift register is then pulsed 15 times with an internal signal, in slightly less than 90 ms. This results in a train of pulses that represent the state of the various 15 stages of the binary counter. This train of pulses is called a binary word and indicates the intensity of the light at the exit slit during the time the gate was open.

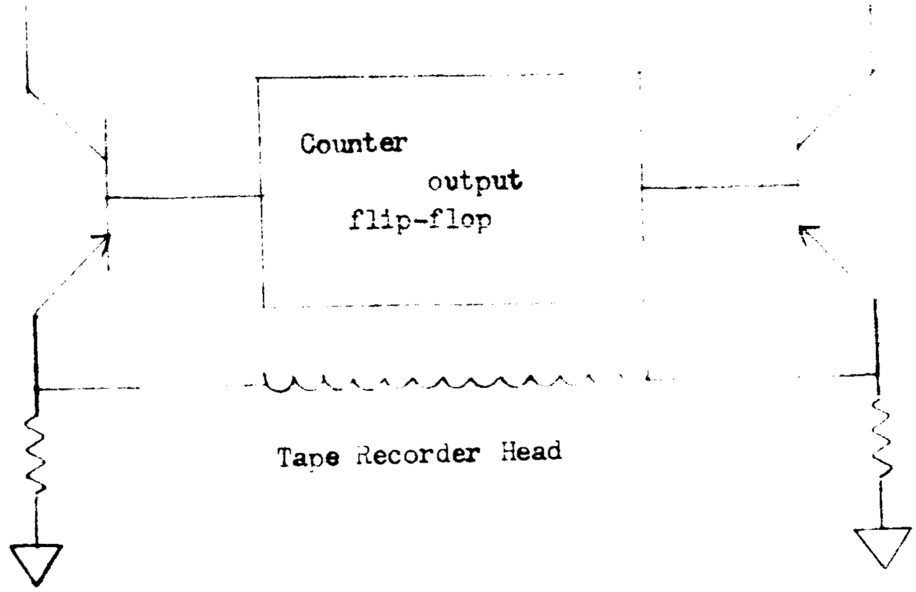
7. TAPE RECORDER AND TELEMETRY REQUIREMENTS

The telemetry and probably the tape recorder will be provided by Ball Brothers. The specifications for this equipment are set down in this section.

7.1 TAPE RECORDER

Since the information to be stored is in digital form, saturation recording would be most desirable. No record electronics would be required since the output of the binary counter could be used directly to drive the recorder heads. This might require a special low-impedance head winding but the output of the counter will have sufficient power. To provide both positive and negative saturation recording, the head could be connected between the emitters of emitter-followers which could be connected to the counter output flip-flop, as shown in Fig. 7. The current would then flow one way or the other through the head winding, depending on which transistor was conducting. This would give saturation recording in both directions.

This type of recording would eliminate a bias and erase oscillator as well as high specification on wow and flutter. This recording process would eliminate many of the adjustments that are necessary on a normal FM or AM recorder. The output of the binary counter is a NRZ (Non Return to Zero) system. This means that if several "1"s appear in sequence at the counter output, the voltage output would be constant, i.e., it would not return to zero between digits. This results in a constant flux on the tape during these periods.



Tape recorder head drive which will allow positive and negative saturation recording.

Figure 7

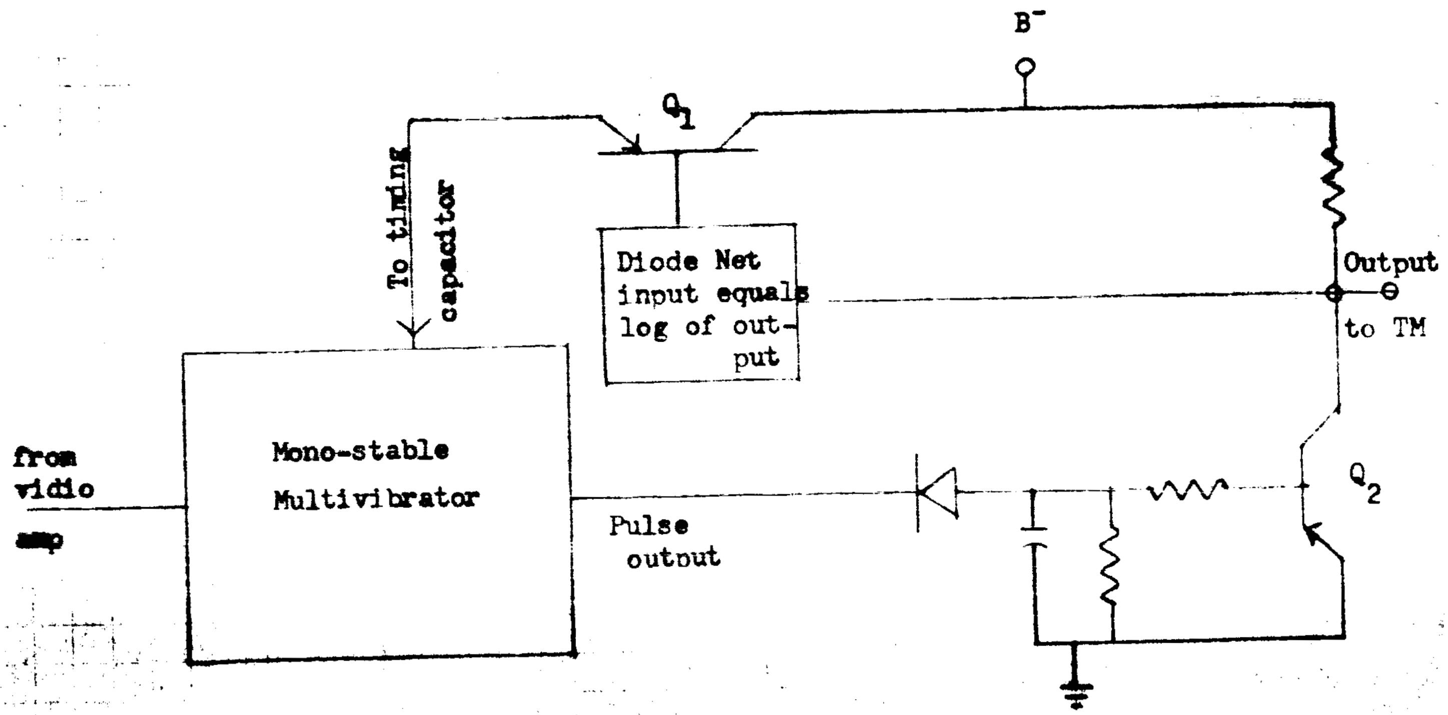


Figure 8

The output, since it is the derivative of the input, is a pulse, the polarity of which indicates whether the change is from "1"s to "0"s or from "0"s to "1"s. These pulses will have a width that approaches that allotted to one digit when the packing density approaches the maximum, i.e., a pulse width of 90/15 ms. These pulses will drive an output flip-flop, which will in turn drive the SCO (Sub-Carrier Oscillator).

Storage for two-thirds of one 90-minute orbit will require slightly over 0.5×10^6 bits per spectrometer. If a tape recorder with 0.5-mil heads is used, and if the tape travels 2 gap widths per bit (which should be possible but is near the limit) then it is possible to store 1000 bits per inch of tape. At this packing density, slightly over 500 inches or 45 feet of tape would be required. It is assumed that two tracks will be used, one for each spectrometer. In order to achieve an effective "gap on the tape" of 0.5 mil, the holdback tensions become quite critical to adjustment. Therefore, it would be advisable to use at least 90 feet of tape to improve the reliability to a safe figure.

7.2 TELEMETRY

It is desired that 3 telemetry channels be used per spectrometer. Channel 1 would be used for a quick look at the output data while the satellite is in contact with the ground station. This information would not be stored. The output would be analog with a large dynamic range in spectral intensity. No great accuracy would be needed; therefore, a circuit such as shown in Fig. 8 could be used to accomplish this purpose. This output at the ground station could be fed directly into a chart recorder and would give an intensity vs. wavelength plot with the intensity plotted logarithmically.

In the circuit of Fig. 8, the monostable multivibrator provides a constant level output pulse (regardless of input pulse size above the triggering level). The width of this pulse is varied by the feedback loop. As the pulse rate increases, the collector voltage of Q_2 becomes less negative, thereby causing more current to flow in Q_1. This current decreases the charge time of the monostable multivibrator and thus the output pulse length is reduced. This of course results in less output voltage across the integrating network. The diode network in the feedback loop provides the non-linearity to give the logarithmic output. The design of this network is such that the log of the voltage across the network is proportional to the current input (because of the large value of R_1).

Channel 2 would transmit the digital information from the binary counter directly to the ground while the satellite is in contact with the ground station. Thus channel 2 would transmit the information that would otherwise be lost while the tape recorder was in the play-back mode. Also some data could be obtained from the satellite even if the tape recorder and its associated components failed. This information could be sent through a channel of about 300-cps bandwidth per spectrometer.

Channel 3 would handle the data stored in the tape recorder. This requires the greatest bandwidth of the 3 channels. The 0.5×10^6 bits per spectrometer must be transferred to the ground in a maximum of 5 minutes. The bit rate is therefore 1700 bits per sec per spectrometer. This bit rate would require at least a 3.4-kc channel per spectrometer.

This is very high for an FM/FM system. The total bit rate of the first S-16 satellite is just over 20 bits per sec real time in comparison to more than 300 bits per sec for the pointed experiment on the system proposed here. The first satellite, however, sends analog data to the ground. This analog data must be of high precision. For example, in the University of Colorado experiment, the rotational position of the grating (which indicates wavelength) is being transmitted by pulse amplitude modulation on the FM/FM link. They will determine this position with a resolution of 1 in 300 (3-A coverage with .01-A resolution).

For this resolution, high dynamic range and very good pulse shape are required. Since the system proposed in this report is a digital one, dynamic range need not be over a few db and the pulse shape can be half a sine wave. This relaxation in telemetry link requirements can result in an increase in bandwidth.

It is therefore proposed that the satellite transmitter be modulated with an SCO. However, instead of the frequency of the SCO being varied as in an FM system, the SCO will be turned off and on in response to "1"s or "0"s of the binary code. This would require a bandwidth into the transmitter of 3.4 kc if the lower SB of the SCO were filtered.

The telemetry transmitter uses standard telemetry channels, the highest channel having an SCO of 22 kc. The bandwidths of these channels are ± 7.5 percent. The upper 4 channels have SCO frequencies of 7.35 kc, 10.5 kc, 14.5 kc and 22 kc. The bandwidth from the lower edge of the 14.5-kc channel to the upper edge of the 22-kc is 13.4 DC, which is the bandwidth into the transmitter that these two channels occupy. If the use of these two channels for the spectrometers can be justified, the bandwidth is adequate for the large bit-rate required by the proposed system. If two more channels can be used for the spectrometer experiment, another 6.0 kc will be available. This additional bandwidth would give a 19.4-kc channel, thus allowing sidebands of the SCO to be fed to the transmitter (6.8-kc + 6.8-kc + 5.8-kc channel separation). This, of course, would simplify the ground equipment.

Appendix 2

MECHANICAL AND OPTICAL DESIGN CONSIDERATIONS

Preliminary design studies have been carried out on two scanning spectrometers, which are to be integrated into a single package. Schematic diagrams of the two instruments are shown in Figs. 1a and 1b. The first spectrometer operates in the wavelength region λ 1500 to λ 500 and the second in the range of λ 600 to λ 75. The long wavelength spectrometer is of the Johnson-Onaka type, in which the grating, G, rotates about a point, C, located so as to produce minimum aberration over the desired spectral range (see Fig. 1a). The spherical collector mirror, M (focal length = 19 cm), produces an image of the sun 1.7 mm in diameter on one of two available entrance slits, N. In angular measure the larger slit is approximately 1 minute of arc long and 0.7 minute of arc wide (60 μ x 42 μ) at the focus of the collector mirror. As focused by the 1200 line/mm concave grating, G, which has a radius of curvature of 70 cm, the slit width is 0.5 A. The smaller slit is also one minute of arc in length but only about 0.1 A (10 μ) wide. The exit slit, X, is identical in size and shape to the narrow entrance slit. Two rates of spectral scan speed are available upon command from the ground. The slow scan requires twenty minutes for a complete sweep from 500 A to 1500 A and is only used with the narrow entrance slit; a complete fast scan takes just four minutes and is used only with a wide slit. The latter combination will be used to observe solar regions of high activity, such as plage areas where flares are expected.

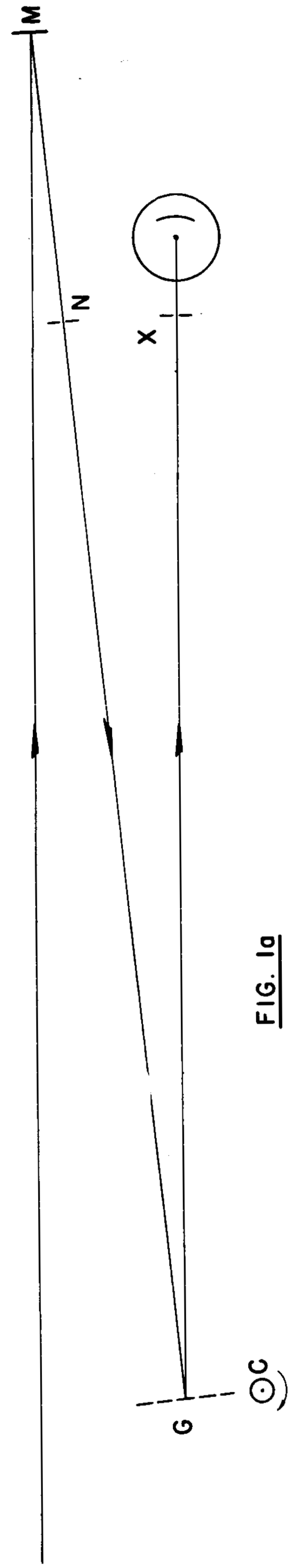


FIG. 1a

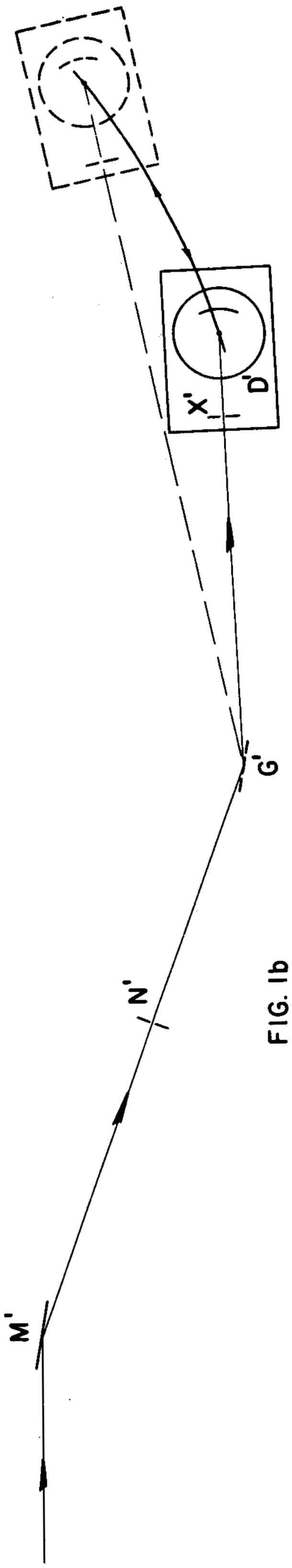


FIG. 1b

The short wavelength spectrometer is of the conventional grazing-incidence type (see Fig. 1b) with the detector, D', and the exit slit, X', moving along the Rowland circle. The collector mirror, M', is to have a toroidal shape approximating a far off-axis paraboloid and will produce a solar image with a diameter of 1.7 mm on the entrance slit, N'. Two entrance slits coupled with scan speeds of four and twenty minutes are available. The larger entrance slit will be approximately 2 minutes of arc (120 μ) on a side; the narrower entrance slit and the exit slit will be of the same length but 25 μ wide. The concave grating, G', is ruled at 1200 lines/mm and has a radius of curvature of one meter; the corresponding spectral band passes of the two slits are approximately 1 A and 0.2 A.

For detectors, both spectrometers will use Bendix windowless photomultipliers with tungsten photocathodes, which have zero response to wavelengths longer than 1500 A and a relatively flat response to shorter wavelengths.

OPTICAL DESIGN

Traverse of the spectral range 500 A to 1500 A in the long-wave spectrometer is achieved by rotating the grating about an offset pivot. The design is based upon achieving perfect focus at 500 A, i.e., the entrance and exit slits are on the Rowland circle for this wavelength. The offset pivot on the grating provides a certain amount of correction for the defocusing caused by the rotation. The distance of the pivot arm was chosen to minimize the degree of defocusing over the entire spectral range between 500 A and 1500 A. Detailed calculations have been carried out to 8 significant figures for this purpose, as well as for the exact location of the various optical parts.

The optical design of the short-wave spectrometer is based on an article by Haas and Tousey (Journal of the Optical Society of America, June 1959). Detailed calculations carried to 8 significant figures define the relative locations of the various optical components.

ADJUSTMENT MECHANISMS

Because of the very small tolerances required for optical alignment, adjustment mechanisms are provided throughout the spectrometers for alignment of the various optical points. At the time of manufacture, all optical points should be aligned within one or two thousandths of an inch of the theoretical position. To achieve final alignment, adjustment mechanisms have been applied to all critical parts of the spectrometers. Most of these consist of a dovetail arrangement on a screw. In many cases, three such adjustments have been provided in one mechanism, allowing adjustments to be made in three dimensions. The motions of the dovetail adjusters are achieved by differential screws, which will give approximately 6 ten-thousandths of an inch of motion for one revolution of the screw. Provisions have also been made for coarse adjustments.

DRIVING MECHANISMS

Rotation of the grating in the long-wave spectrometer and translational motion of the photomultiplier in the short-wave spectrometer are accomplished by Sigma Cyclonome stepping motors driving through a reduction system of friction wheels. Approximately 6000 steps are required both to rotate the long-wave grating through approximately 3.5 degrees and to move the short-wave exit slit through its range of about 8 inches along the Rowland circle. The long-wave grating is pivoted about an elastic hinge about the theoretical point of rotation, and driven by a cam on the rear of the grating. The 500-A wavelength will appear on the exit slit when the follower is at the minimum radius of the cam. No provision has been made for the quick return of the long-wave grating. Consequently, as the satellite comes out of the shadow of the sun, the long-wave spectrometer will begin scanning at the wavelength position it had when the satellite entered the shadow of the earth. However, since two full sweeps and one partial sweep of the spectrum are possible during any one orbit, there should be no limitation on the spectral range observed in each orbit. For the short-wave spectrometer, a quick return is provided to bring the carriage back into its initial position. As requested by Ball Brothers, the quick return locates the center of gravity of the package at the geometrical center of the package at the time the satellite goes into the shadow of the earth. The stepping-motor friction wheel reducers and some of the electronic amplification apparatus move together on the same carriage with the exit slit.

SLIT CHANGERS

The exit slits for both the long- and short-wave spectrometers will be changed during flight. Different slit-changing mechanisms will be used for each of the two spectrometers. In the case of the long-wave spectrometer, a rotary device will rotate through an angle of 90 degrees to present first the narrow slit, and then the wide slit at the proper optical position. The rotation will be accomplished by the selective positioning of permanent magnets and electromagnets.

ADJUSTMENT OF SLITS

The slit plates are to be ground so that their edges are straight to within 10 millionths of an inch. Final adjustment is to be made by the calibrator. It is suggested that the slits be set under a microscope and then assembled on their respective holders into position on the tables or the slit changer, as the case may be. The positioning holes for the holders should position the slits within a thousandth of an inch from the true theoretical optical position.

LAUNCH-LOCKING MECHANISMS

A mechanism is being provided to lock the grating of the long-wave spectrometer in position during launch. The mechanism consists of a pin inserted into the side of the grating mount, and released upon an electrical impulse to a solenoid, which in turn releases the trigger. A similar device is provided for the carriage on the short-wave spectrometer.

AEROBEE ROCKET SPECTROMETER

Preliminary designs have also been carried out for a long-wave spectrometer to be flown in an Aerobee Rocket. The optical parts in this package are identical with those designed for the satellite instrument. Most of the other parts in the package are similar to those in the satellite instrument, with the exception of details necessary for attachment. The main exception is that the driving mechanism is no longer a system of friction wheels, but a system of gears. However, the gears are driven by the same stepping motor and the output is also to a cam, although the cam is of different configuration than that of the satellite design. The motion of the grating is confined to about one-third of the spectral region traversed in the long-wave spectrometer of the satellite package. The rate of the stepping motor is designed to give three full traversals of the spectrum during the time that the rocket is above the atmosphere. The optical axis of the spectrometer was chosen to be 14 degrees from the axis of symmetry of the rocket in order that the spectrometer should have the highest probability of having the sun in view throughout its flight. Provision has been made at the base of the spectrometer for the attachment of the eye-blocks of the pointing eyes to be provided by Ball Brothers. Adjustment has been provided for the entrance and exit slits, the object mirror, and one side of the grating hinge. Further, a gross alignment of the supports for the gear drive is possible. This alignment provides the calibrator with freedom to choose any given spectral range.

THERMAL BALANCE

Theoretical calculations have been carried out on the transfer of energy from outer space from the sun, to and from the earth, to and from other parts of the satellite package. Equations have been set up that consider the variations in the thermal absorption and emission during various orientations of the satellite with respect to the earth. Some calculations have been performed considering various orbital heights and eccentricities, and seasonal variations. However, final calculations on the thermal balance of the satellite package have not yet been completed.